Input Impedance Conditions for Minimizing the Noise Figure of an Analog Optical Link

Edward I. Ackerman, *Member, IEEE,* Charles Cox III, *Senior Member, IEEE,* Gary Betts, *Member, IEEE,* Harold Roussell, Kevin Ray, and Fred O'Donnell

***Abstract*— We use an equivalent-circuit model to derive the minimum noise figure (NF) for an amplifierless optical analog link with a perfect lossless input impedance-matching circuit. This minimum, which is 3 dB, is called the *lossless passive match limit*. We contrast the link model with an analogous equivalent-circuit model for a transistor amplifier, which does not have the same limiting NF under this impedance-matching condition. It turns out that the link's 3-dB limit arises from the ohmic nature of the impedance of the device that modulates the light, and not from the electro-optic conversion processes in the link. In a prior experimental link with near-perfect impedance matching, dissipative loss in our input circuit precluded achieving a measured NF of less than 4 dB. Investigation of the effects of input impedance mismatch indicates that mismatch can actually lower the NF to below 3 dB even when dissipative loss is present in the input circuit. We have used this mismatch effect to reduce the measured NF of our link to 2.5 dB at 130 MHz. We believe this is the first demonstration of amplifierless-link NF of less than 3 dB.**

***Index Terms*— Impedance matching, modeling, noise, optical-fiber communication, optical modulation/demodulation.**

I. Introduction

MINIMIZATION of analog optical link noise figure (NF) is very important in applications such as remote sensing and receive antenna remoting. It is usually accomplished by using a low-noise amplifier before the modulation device (i.e., the external modulator or directly modulated semiconductor laser). However, relying on large preamplifier gain to counteract a large-link NF can yield a significantly smaller overall dynamic range. Therefore, it is important to understand how the *amplifierless* link's NF can be minimized.

To date, the lowest reported link NF's have been demonstrated at relatively low frequencies (usually below 1 GHz) [1]–[4] because, at lower frequencies, it is easier to obtain lasers, modulators, and detectors with high-efficiency and low-noise characteristics, and it is easier to achieve high gain from impedance matching. As device technology improves, it should be possible to extend these results to higher microwave frequencies. Additionally, at the same frequencies where the lowest NF results have been reported, higher efficiency and lower noise devices should enable the NF to be decreased even further. However, there is a limit to how low the NF of a link can be. Specifically, it has been previously claimed [5]–[7] that, in the absence of electronic preamplification, an optical analog link with perfect lossless input impedance matching has a minimum NF of 3 dB. This *lossless passive match limit*, as we call it, has generated confusion because: 1) it is widely known that other active two-port electronic devices, such as transistor-based low-noise amplifiers (LNA's), have exhibited microwave NF's below 3 dB even when impedance matched for maximum gain rather than for minimum NF and 2) it is not widely known in what specific ways the analog optical link differs from these devices and why the lower limit to its NF is, therefore, different.

In this paper, we present a derivation of the lossless passive match limit, employing an equivalent-circuit model for an analog optical link with a lossless input impedance-matching circuit. We then derive how excess loss in the input impedance-matching circuit affects the NF. We find that, as one would expect, this loss at the link input raises the noise-figure limit from 3 dB to a value equal to 3 dB, plus the impedance-matching circuit loss in decibels. This effect does not bode well for attempts to achieve very low amplifierless-link NF's. However, as we discuss further on, we know that there exists at least two techniques for reducing the amplifierless-link NF to below 3 dB, even in the presence of input circuit loss. We have chosen to experimentally pursue one of these techniques, which involves the use of an interface circuit that transforms the modulator impedance but purposely refrains from matching it to the source impedance.

II. Derivation of the Passive Match Limits

An NF is defined as "the ratio of the available output noise power per unit bandwidth to the portion of that noise caused by the actual source connected to the input terminals of the device, measured at a standard temperature of 290 K" [8]. That is,

$$\mathrm{NF} \equiv 10 \log \left[\frac{\overline{N}_{\mathrm{out}}}{\overline{N}_{\mathrm{in}} G}\right] \tag{1}$$

where $\overline{N}_{\mathrm{out}} \equiv \overline{N}_{\mathrm{in}}G + \overline{N}_{\mathrm{added}}$, $\overline{N}_{\mathrm{in}} \equiv kT$, G is the available gain, k is Boltzmann's constant, and $T \equiv 290$ K. Note that determination of link NF using (1) requires knowledge of only two quantities: the available gain and noise added by the link. In this paper, we derive G and $\overline{N}_{\mathrm{added}}$ for an amplifierless intensity-modulation/direct-detection optical

Manuscript received November 9, 1997; revised May 28, 1998. This work was supported under U.S. Air Force Contract F19628-95-C-0002.

The authors are with the MIT Lincoln Laboratory, Lexington, MA 02173 USA.

Publisher Item Identifier S 0018-9480(98)09045-0.

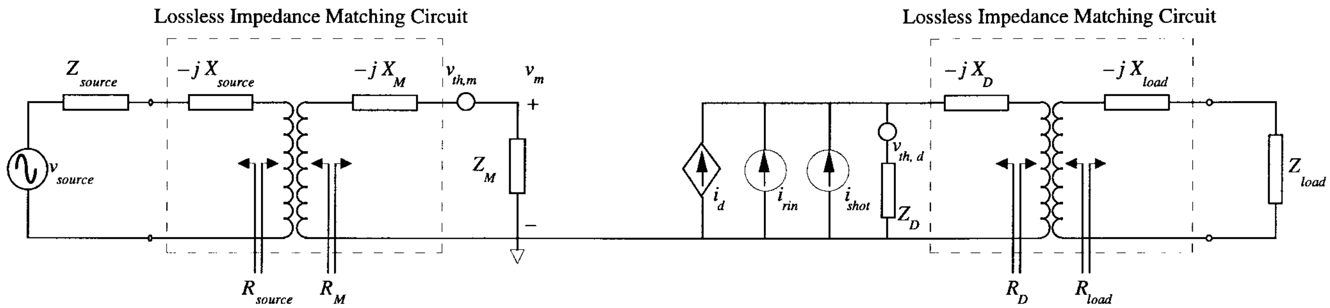


Fig. 1. Equivalent circuit of an intensity modulation/direct-detection optical link with perfect lossless input and output impedance-matching circuits.

link under a number of modulator interface-circuit conditions, beginning with the case of a lossless circuit that perfectly impedance matches the modulator to the RF source. Except where we specify, "modulator" can be interpreted to mean either an external electrooptic intensity modulator or a directly intensity-modulated semiconductor laser.

A. Link with Lossless Impedance Matching

Fig. 1 shows an equivalent circuit for an intensity-modulation/direct-detection optical link with perfect and lossless modulator and detector impedance-matching circuits. At any frequency, one can model the modulator and detector, respectively, as impedances Z_M and Z_D, whose real parts R_M and R_D represent physical ohmic resistances with associated thermal noise. Similarly, Z_{source} and Z_{load} represent the RF source and load impedances, respectively, and R_{source} and R_{load} are their real parts. Reactive circuit elements transform Z_M and Z_D so that the link input and output impedances are Z^*_{source} and Z^*_{load}, respectively.

The optical link between the modulator and detector is represented in Fig. 1 by a voltage-dependent current source. The photocurrent is a function of the optical power incident on the photodetector, which is itself a function of the voltage across the modulator. The link's transducer gain g_1 can be expressed as the ratio of the small-signal photocurrent i_d to the small-signal voltage v_m across the modulator electrodes, and has a dimension of amperes/volt. The magnitude of g_1 is dictated by the type of modulator, its bias condition, and the total optical loss in the link. Thus, one derives the link's available gain G as follows (we use "gain" here in the most general sense; i.e., insertion loss is simply less than unity gain):

$$G \equiv \frac{P_{\text{out}}}{P_{\text{in, av}}} = |g_1|^2 \frac{|Z_M|^2 |Z_D|^2}{4R_M R_D}. \tag{2}$$

The equivalent circuit also includes two voltage sources and two current sources representing noise that is always generated in any intensity-modulation/direct-detection link: $v_{\text{th},\,m}$ and $v_{\text{th},\,d}$ are the thermal noise voltages generated by the ohmic resistive portions of the modulator and detector impedances, respectively; i^2_{rin} and i^2_{shot} are the current spectral densities of the photodetected optical relative intensity noise (RIN) and shot noise, respectively. From the equivalent-circuit representations of these four noise sources, one can calculate the output noise per unit bandwidth added by the link

$$\overline{N}_{\text{added by link}} = \frac{|v_{\text{th},\,m}|^2}{4R_M} |g_1|^2 \frac{|Z_M|^2 |Z_D|^2}{4R_M R_D} + \frac{|v_{\text{th},\,d}|^2}{4R_D} + i^2_{\text{rin}} \frac{|Z_D|^2}{4R_D} + i^2_{\text{shot}} \frac{|Z_D|^2}{2R_D}. \tag{3}$$

Notice in (3) that because the four noise sources are uncorrelated, the total noise they contribute at the link output is determined by summing the noise powers—each of which is calculated from the square of a voltage or current—rather than by summing voltages or currents and then squaring.

In this paper, we continue to represent the effects of laser RIN and shot noise using the very simple forms shown in (3). More detailed expressions for these two noise powers are given elsewhere [5]–[7]. The thermal-noise terms are more relevant to our discussion, since they give rise to the lower limits to NF. The thermal-noise-voltage spectral densities in (3) are calculated using the following relationships [9]:

$$|v_{\text{th},\,m}|^2 = 4kTR_M \tag{4}$$

$$|v_{\text{th},\,d}|^2 = 4kTR_D. \tag{5}$$

Substituting these into (3) yields the noise per unit bandwidth added by the link

$$\begin{aligned}\overline{N}_{\text{added by link}} &= kT|g_1|^2 \frac{|Z_M|^2|Z_D|^2}{4R_M R_D} + kT + (i^2_{\text{rin}} + i^2_{\text{shot}}) \frac{|Z_D|^2}{4R_D} \\ &= kTG + kT + (i^2_{\text{rin}} + i^2_{\text{shot}}) \frac{|Z_D|^2}{4R_D}.\end{aligned} \tag{6}$$

Therefore,

$$\begin{aligned}\overline{N}_{\text{out}} &\equiv kTG + \overline{N}_{\text{added by link}} \\ &= 2kTG + kT + (i^2_{\text{rin}} + i^2_{\text{shot}}) \frac{|Z_D|^2}{4R_D}\end{aligned} \tag{7}$$

and the NF for an amplifierless optical analog link with perfect lossless input impedance matching is calculated by substituting (7) into (1)

$$\text{NF} = 10 \log \left[2 + \frac{1}{G} + \frac{(i^2_{\text{rin}} + i^2_{\text{shot}})|Z_D|^2}{4R_D kTG} \right]. \tag{8}$$

For large G, (8) reduces to the *lossless passive match limit* of $\text{NF} = 10 \log [2]$, or 3 dB.

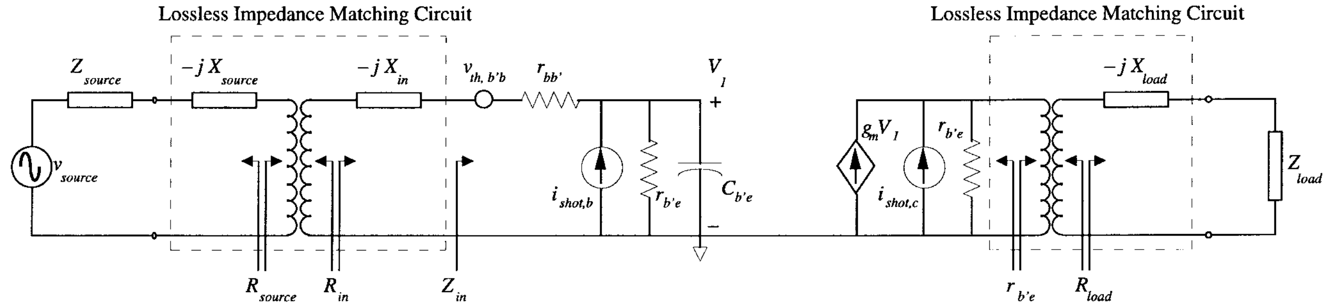


Fig. 2. Hybrid-π small-signal equivalent-circuit model of a BJT transistor (after [10]).

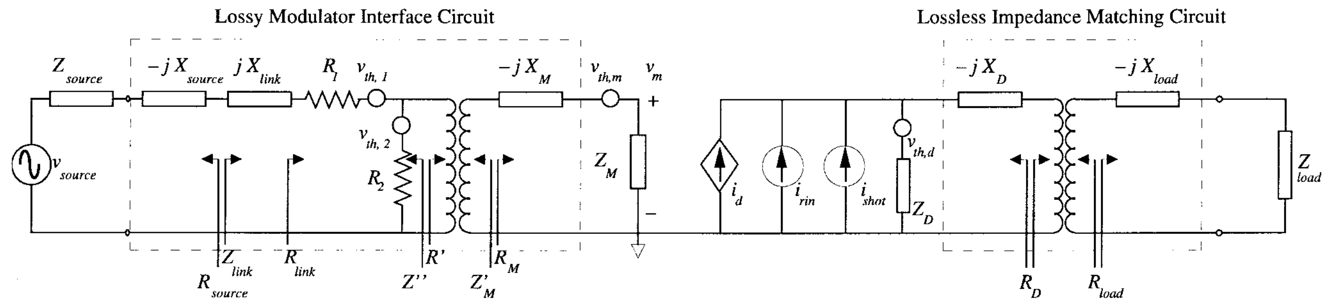


Fig. 3. Equivalent circuit for an intensity-modulation/direct-detection optical link with arbitrary input impedance Z_{link}, and resistors R_1 and R_2 to simulate loss in the modulator interface circuit.

B. Bipolar Transistor with Lossless Impedance Matching

The noise-figure derivation has yielded a 3-dB lossless passive match limit for any two-port three-terminal active device in which the entire real part of the input impedance is ohmic. Since intensity-modulation optical links fall into this category, they differ from many other active two-ports. For instance, most transistors, including bipolar junction transistors (BJT's) and MESFET's, have input impedances with real parts arising not entirely from physical ohmic resistances. In the hybrid-π small-signal equivalent circuit of a bipolar junction transistor, which is shown in Fig. 2 with lossless input and output impedance-matching circuits, the elements $r_{b'e}$ and $C_{b'e}$ represent the base-to-emitter impedance [10]. These elements yield the same voltage-to-current relationship at the base that arises from the base-to-emitter characteristic, but their impedance is an incremental or effective one, rather than an ohmic one. Consequently, whereas the base-spreading resistance $r_{bb'}$ contributes thermal noise due to its ohmic nature, the element $r_{b'e}$ does not. If, for simplicity's sake, we neglect shot noise and consider only low frequencies (where the effect of $C_{b'e}$ is negligible), we can derive the NF in the same way as we did for the optical link

$$\overline{N}_{\text{added by BJT}} = \frac{r_{bb'}}{r_{bb'} + r_{b'e}} kTG + kT \tag{9}$$

$$\begin{aligned}\overline{N}_{\text{out}} &\equiv kTG + \overline{N}_{\text{added by BJT}} \\ &= \left(1 + \frac{r_{bb'}}{r_{bb'} + r_{b'e}}\right) kTG + kT \end{aligned} \tag{10}$$

and

$$\text{NF} = 10 \log \left[1 + \frac{r_{b'b}}{r_{bb'} + r_{b'e}} + \frac{1}{G}\right]. \tag{11}$$

Note that even in this impedance-matched condition, which seldom yields the minimum NF for a transistor amplifier, NF can be less than 3 dB. To minimize NF, bipolar transistor designers seek to minimize the base-spreading resistance $r_{bb'}$.

C. Link with Lossy Modulator Impedance-Matching Circuit

Fig. 3 shows the equivalent circuit for an intensity-modulation/direct-detection optical link, as was shown in Fig. 1, but with some important changes. Instead of transforming R_M (the real part of the modulator impedance) perfectly and losslessly to the source resistance R_{source}, the lossless transformer makes the modulator appear to have a resistance R'. The loss of the matching circuit is then simulated by adding a combination of series and parallel resistances (R_1 and R_2, respectively), resulting in a purely real link input impedance R_{link}. An additional component with reactance X_{link} is shown in series with this configuration because, in Section III, we will be examining NF as a function of complex link input impedance. For now, however, we are still concerned with the impedance-matched case, i.e., where $X_{\text{link}} = 0$, $R_{\text{link}} = R_{\text{source}}$, and $Z'' = R'$.

To see most clearly how NF varies with impedance-matching circuit loss, it is desirable to express the resistances R_1 and R_2 as a function of the loss they impose. This can be done by invoking several relationships between the quantities shown in Fig. 3, beginning with

$$R_{\text{link}} = R_1 + (R_2 || R') = R_1 + \frac{R_2 R'}{R_2 + R'} \tag{12}$$

or, equivalently,

$$R' = \frac{R_2 R_{\text{link}} - R_1 R_2}{R_1 + R_2 - R_{\text{link}}}. \tag{13}$$

Notice also from Fig. 3 that

$$Z'' = R_2 \| (R_{\text{link}} + R_1) = \frac{R_{\text{link}} R_2 + R_1 R_2}{R_{\text{link}} + R_1 + R_2} \tag{14}$$

from which we can conclude that

$$R_2 = \frac{R_{\text{link}}^2 - R_1^2}{R_1} \tag{15}$$

and

$$R' = \frac{R_{\text{link}}^2 - R_1^2}{R_{\text{link}}}. \tag{16}$$

At this point, the gain derivation that resulted in (2) for the lossless impedance-matching case (which we will now call G_{lossless}) can be repeated to yield an expression for link gain that includes the effects of circuit loss

$$G = |g_1|^2 \frac{|Z_M|^2 |Z_D|^2}{4R_M R_D} \frac{R_{\text{link}} - R_1}{R_{\text{link}} + R_1} = \frac{R_{\text{link}} - R_1}{R_{\text{link}} + R_1} G_{\text{lossless}^*}. \tag{17}$$

Defining $G \equiv G_M G_{\text{lossless}}$, where G_M is the circuit's excess gain (less than one for a passive circuit), it is clear that the circuit loss must be related to resistance R_1 as follows:

$$G_M = \frac{R_{\text{link}} - R_1}{R_{\text{link}} + R_1} \tag{18}$$

and, therefore,

$$R_1 = \frac{1 - G_M}{1 + G_M} R_{\text{link}} \tag{19}$$

$$R_2 = \frac{4G_M}{1 - G_M^2} R_{\text{link}} \tag{20}$$

and

$$R' = \frac{4G_M}{(1 + G_M)^2} R_{\text{link}}. \tag{21}$$

These equations make sense in that removing the lossy elements from Fig. 3 results in $G_M = 1$ (so that $G = G_{\text{lossless}}$), which is equivalent to setting $R_1 = 0$ and $R_2 = \infty$.

With the resistances R_1 and R_2 now expressed as functions of the loss they represent, it is possible to derive the NF as a function of that loss as well. Fig. 3 shows thermal noise voltages generated by these two impedances. Including these two additional noise voltages in the derivation yields the following expression for the NF of a link with perfect, but not lossless input impedance matching

$$\begin{aligned} \text{NF} &= 10 \log \Bigg[1 + \frac{1 - G_M}{1 + G_M} + \frac{1}{G_M} \frac{1 - G_M}{1 + G_M} + \frac{1}{G_M} + \frac{1}{G} \\ &\qquad + \frac{(i_{\text{rin}}^2 + i_{\text{shot}}^2)|Z_D|^2}{4R_{\text{D}} kTG} \Bigg] \\ &= 10 \log \left[\frac{2}{G_M} + \frac{1}{G} + \frac{(i_{\text{rin}}^2 + i_{\text{shot}}^2)|Z_D|^2}{4R_D kTG} \right]. \end{aligned} \tag{22}$$

For very large G, the general passive match limit to NF is, therefore, $10 \log[2/G_M]$, which is equal to 3 dB plus the loss in the matching circuit (recall that G_M is the excess gain—which is ≤ 1 for passive circuits—and, therefore, $1/G_M$ is the excess loss). This result is what one expects when adding passive attenuation to the front end of a network.

The general passive match limit helps to explain the measured amplifierless link G and NF we have reported previously [4]. We had impedance matched the low-V_π external modulator in a narrow-band link to the 50-Ω RF source using a circuit for which G_M had been independently measured to be -0.7 dB. When we used a very large optical power (400 mW) at the input to the modulator, we measured $G = 26.5$ dB and NF = 4.2 dB at 150 MHz. Using $G = 26.5$ dB along with $G_M = -0.7$ dB and the other parameters in our link model resulted in a predicted NF of 4.0 dB, which we interpret as confirmation of the model. Since achieving this result, we have attempted to reduce the modulator interface-circuit loss to reduce the NF. We have also sought a means of reducing the NF even further by somehow circumventing the lossless passive match limit of 3 dB, as is discussed in Section III.

III. Methods for Reducing NF to Below 3 dB

In Section II, we showed that if the modulator in an amplifierless link is perfectly matched to the RF source impedance, then the ohmic resistive portion of the modulator impedance contributes kTG to $\overline{N}_{\text{added by link}}$, so that $\overline{N}_{\text{out}}$ is at least $2kTG$ and NF is at least 3 dB. Reducing NF to less than 3 dB requires a reduction of the modulator impedance's contribution to $\overline{N}_{\text{added by link}}$ so that it is significantly less than kTG. We are currently aware of two methods for accomplishing this, which are: 1) using a traveling-wave electrooptic modulator, in which ideally the only added thermal noise on the electrodes is generated by the termination impedance (which has a diminished contribution to the output noise because it is launched in the "counterpropagating" direction [11]) and 2) removing the "match" constraint by using any type of modulator, but purposely mismatching it to the RF source impedance in such a way that the thermal noise voltage its ohmic impedance generates is divided so that less of it is ends up being across the modulator. Both of these methods for potentially reducing the NF can result in less than 3-dB NF, even in the presence of some matching-circuit loss, as is explained further in this section.

A. Using a Traveling-Wave Modulator

If a traveling-wave modulator is used, then, in the lossless match case, the electrode characteristic impedance and electrode termination impedance are both equal to R_{source}, and the only contribution to $\overline{N}_{\text{added by link}}$ from the modulator is the thermal noise kT generated by the termination resistance R_{source}. This thermal noise is launched onto the traveling-wave electrode from the end opposite from the end onto which the input signal is launched. In a traveling-wave modulator, the electrical and optical phase velocities are matched in the direction that the signal is launched, whereas the thermal noise from the termination resistance propagates in the direction counter to this preferred launch direction. Therefore, at any modulation frequency where the signal's effective wavelength in the modulator material is much less than the traveling-wave electrode length, the modulator thermal noise will contribute negligibly to $\overline{N}_{\text{added by link}}$, such that the lossless passive

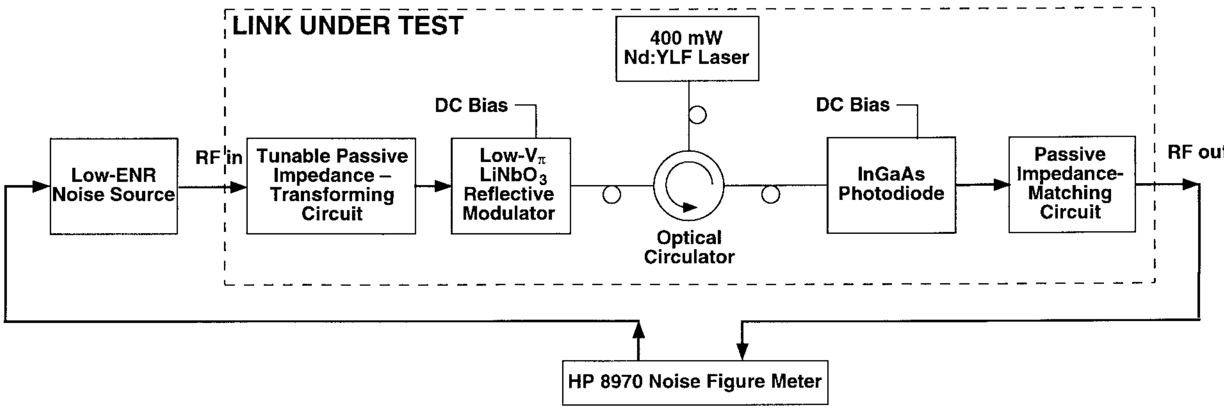


Fig. 4. Setup for measuring the NF of the experimental amplifierless link with tunable input impedance.

match limit for this type of modulator is close to 0 dB (depending on the modulation frequency) [11].

We have chosen not to attempt this method of obtaining less than 3-dB NF because it is easier to minimize the shot noise contribution to $\overline{N}_{\text{added by link}}$ (which must be done if very low NF's are to be obtained) using a modulator with a high-impedance lumped-element electrode than it is using a traveling-wave modulator with the same V_π.

B. Removing the "Match" Constraint

It is possible to purposely mismatch the modulator and RF source impedances in such a way that voltage division causes less of the thermal noise voltage generated by the modulator resistance to be imposed across the modulator electrodes. This, in turn, causes the modulator's thermal noise to contribute less than kTG to $\overline{N}_{\text{added by link}}$, making it a second potential method for reducing the link NF to less than 3 dB.

Deriving the link gain and NF for any link input impedance $Z_{\text{link}}(= R_{\text{link}} + jX_{\text{link}})$, assuming lossless impedance transforming circuitry (i.e., $G_M = 1$), yields

$$G = \frac{R_{\text{link}} R_{\text{source}}}{|Z_{\text{link}} + R_{\text{source}}|^2} |g_1|^2 \frac{|Z_M|^2 |Z_D|^2}{R_M R_D} \tag{23}$$

and

$$\text{NF} = 10 \log \left[1 + \frac{R_M^2 |Z_{\text{link}} + R_{\text{source}}|^2}{R_{\text{link}} R_{\text{source}} |R_M + Z_M'|^2} + \frac{1}{G} + \frac{(i_{\text{rin}}^2 + i_{\text{shot}}^2)|Z_D|^2}{4R_{\text{D}} kTG} \right]. \tag{24}$$

A link NF of less than 3 dB can, therefore, be obtained if the modulator interface circuit is designed to yield a link input impedance that causes the second term in (24) to be less than one (provided that high-Q components are used so that the interface circuit loss is minimized). For this to be successful, the third and fourth terms in (24) must remain small even when mismatch causes G to be smaller than its maximum (i.e., perfect input impedance match) value.

This mismatch method is what we have chosen to pursue in our efforts to demonstrate an amplifierless link with less than 3-dB NF. In our actual modulator impedance-transforming circuit, we expect to have some loss, so it will be helpful to have general expressions for gain and NF that include the effects of this loss in addition to the effects of impedance mismatch. Repeating the gain and NF derivations for $Z_{\text{link}} \neq R_{\text{source}}$ and $G_M \neq 1$ yields

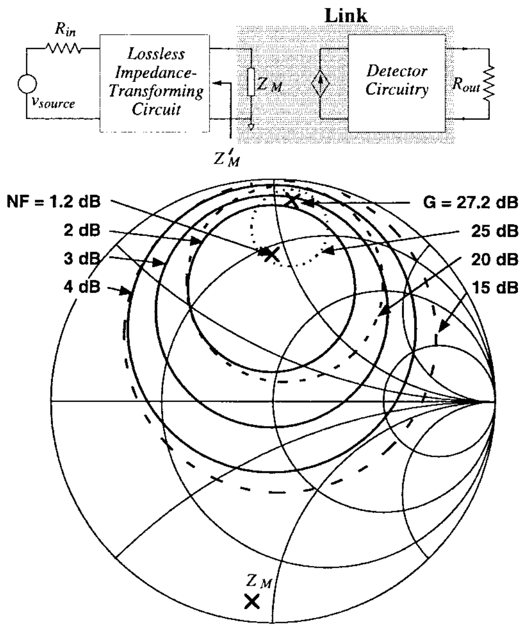
$$G = \frac{R_{\text{link}} R_{\text{source}}}{|Z_{\text{link}} + R_{\text{source}}|^2} |g_1|^2 \frac{|Z_M|^2 |Z_D|^2}{R_M R_D} G_M \tag{25}$$

and (26), shown at the bottom of this page, which are valid for any link input impedance and for any amount of modulator interface-circuit loss. In the case where $X_{\text{link}} = 0$, $Z_{\text{link}} = R_{\text{link}} = R_{\text{source}}$, and $Z_M' = R_M$, these reduce to the expressions already given for the perfect input matching case with loss [(17) and (22), respectively]. When $G_M = 1$, they reduce instead to the equations given for the case where the modulator interface circuit is lossless, but does not impedance match the modulator to the source (23), (24). Finally, combining the lossless and impedance-matching conditions causes (25) and (26) to reduce to (2) and (8), respectively.

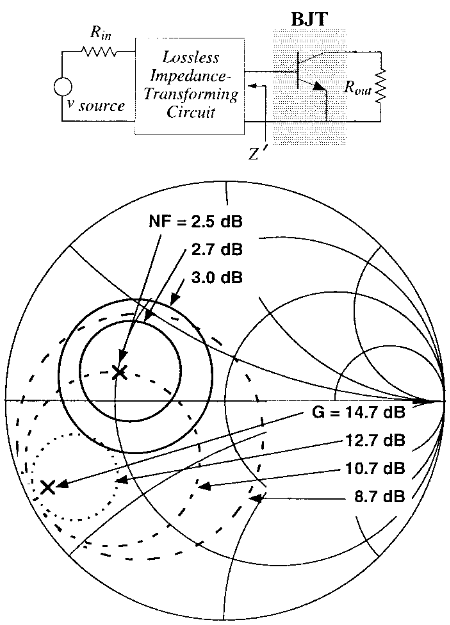
IV. Experimental Results

We assembled an external modulation link using the same laser, modulator, and detector described in [4], but this time we incorporated a modulator interface circuit having adjustable inductance and capacitance values. Varying the tunable reactance values in small increments, we used a calibrated HP8510 network analyzer to measure the resulting link input impedance Z_{link}, and used the experimental setup shown in Fig. 4 to measure the corresponding link NF. We calibrated the HP 8970A NF meter with a noise source having low excess

$$\text{NF} = 10 \log \left[1 + \frac{1 - G_M}{1 + G_M} \frac{R_{\text{link}}}{R_{\text{source}}} + \frac{1 - G_M}{1 + G_M} \frac{|Z_{\text{link}} + R_{\text{source}}|^2}{4G_M R_{\text{link}} R_{\text{source}}} \frac{[(1 - G_M)R_{\text{link}} + (1 + G_M)R_{\text{source}}]^2 + (1 + G_M)^2 X_{\text{link}}^2}{(R_{\text{link}} + R_{\text{source}})^2 + X_{\text{link}}^2} + \frac{R_M^2 |Z_{\text{link}} + R_{\text{source}}|^2}{G_M R_{\text{link}} R_{\text{source}} |R_M + Z_M'|^2} + \frac{1}{G} + \frac{(i_{\text{rin}}^2 + i_{\text{shot}}^2)|Z_D|^2}{4R_{\text{D}} kTG} \right] \tag{26}$$



(a)



(b)

Fig. 5. (a) Analytically determined constant-G and constant-NF circles on a Smith chart plot of the transformed source impedance Z'_M presented to the modulator in the experimental amplifierless link. (b) Analogous constant-G and constant-NF circles for an example bipolar transistor.[1]

noise ratio (ENR) before measuring the link NF. When the tunable reactive elements in the modulator interface circuit were adjusted such that $Z_{\rm link} = 10.6 - j19.3\ \Omega$ at $f = 130$ MHz, we measured an NF of 2.5 dB at this frequency. When tuned in this way for minimum NF at 130 MHz, the NF varied with frequency fairly strongly; we measured $\text{NF} = 3.5$ dB at 115 and 150 MHz.

We then replaced the link-under-test with a variable RF attenuator and adjusted the dial on this device until the meter measured its NF as 2.5 dB. At this dial setting, we measured an insertion loss of -2.5 dB at $f = 130$ MHz using the network

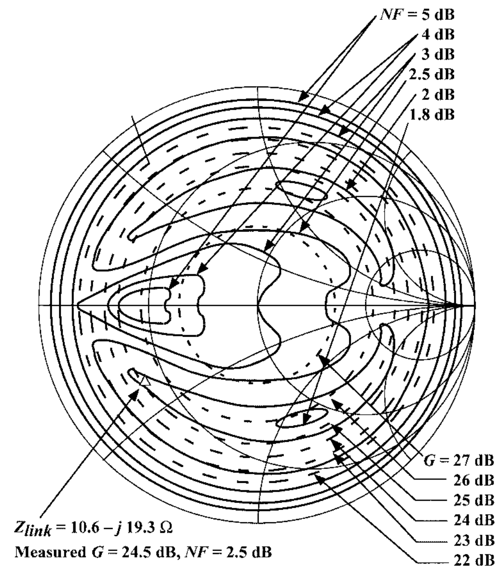


Fig. 6. Conformal mapping of the constant-G and constant-NF circles of Fig. 5(a) from Z'_M-space into $Z_{\rm link}$ space. The impedance where we measured $G = 24.5$ dB and $\text{NF} = 2.5$ dB is also shown.

analyzer. This verified that the meter had given an accurate NF measurement for the link.

To our knowledge, 2.5 dB is the lowest NF ever reported for an amplifierless optical link. The significance of this result is not solely its record-breaking nature, but rather how it adds to our understanding of the relationship between link gain and NF under high-gain conditions. In effect, what we have shown is that when a link has very large available gain, it exhibits many of the same qualities as other high-gain devices such as BJT's and other transistors. Indeed, (23) and (24), which relate G and NF to the source and link input impedances, give rise to constant G and NF loci on the Smith chart in much the same manner that an amplifier's gain and NF behavior was first shown to be represented on a Smith chart plot of its source impedance [12].

Using our model of the experimental link, we have rendered such a Smith chart plot in Fig. 5(a). Note that what is plotted here is not $Z_{\rm link}$, but rather Z'_M, the impedance presented to the modulator by the RF source (which lossless passive circuitry can transform to any finite impedance). Therefore, the effects of matching circuit loss are ignored—just as they always are in similar Smith chart plots published by transistor manufacturers, such as the one shown in Fig. 5(b).[1] Note on the link plot that the lossless passive match limit of 3 dB is represented by the point $Z'_M = Z^*_M$, which is the perfect lossless match case. Two additional facts about Fig. 5 are worth noting. First, unlike the transistor, the link's unidirectionality (i.e., the fact that its S_{12} is zero) causes the perfect match condition to always yield the highest gain, which is why the $G = 27.2$ dB point in Fig. 5(a) is at Z^*_M. Second, if G for our link had not been so large, then the NF and gain contours would more nearly line up with each other; that is, the input matching condition yielding the highest G would also

[1] "A low-noise 4-GHz transistor amplifier using the HXTR-6101 silicon bipolar transistor," Hewlett-Packard Applicat. Note 976, May 1975.

yield the lowest NF, as is the case for most links [as dictated by (23) and (24)].

Due to the difficulty of measuring Z'_M after the input circuit was tuned for lowest NF, we do not know exactly what spot on the Smith chart plot in Fig. 5(a) corresponds to our measured 2.5-dB NF. Instead, we know that at $f = 130$ MHz, our measured result corresponded to $Z_{\rm link} = 10.6 - j19.3\ \Omega$. Using a computer-aided design (CAD) simulation of the tunable modulator interface circuit, we were able to map the circles of constant-G and constant-NF versus Z'_M onto contours (no longer circular) of constant-G and constant-NF versus $Z_{\rm link}$. These are shown in Fig. 6, along with the measured data point of $G = 24.5$ dB and NF $= 2.5$ dB at $Z_{\rm link} = 10.6 - j19.3\ \Omega$. Besides allowing us to compare the analytical and experimental data, Fig. 6 has the additional attractive feature of showing that NF $= 3$ dB at $Z_{\rm link} = Z_{\rm source} = 50\ \Omega$.

V. Conclusions

We have shown why 3 dB is the lowest NF achievable for an amplifierless optical link with perfect lossless impedance matching to the RF source impedance. Investigation into the effects of circuit loss and impedance mismatch on link performance has shown that circuit loss increases the minimum NF, whereas impedance mismatch can actually lower NF to below 3 dB, which we have experimentally demonstrated. Of course, impedance mismatch reduces G, so this technique only yields low NF's when G is so large under the matched condition that $1/G$ is still very small even in the mismatch case. Of additional concern is the effect of input impedance mismatch (which dictates return loss) on the performance of the analog RF system into which the link is inserted.

Acknowledgment

The authors would like to thank P. Haddad, R. Helkey, G. Rezendes, M. Taylor, and A. Yee for technical assistance. The authors' discussions with C. Aitchison, Brunel University, West London, U.K., B. Bridges, California Institute of Technology, Pasadena, and A. Seeds, University College London, London, U.K., increased their understanding of the effects described in this paper.

References

[1] C. H. Cox, III, G. E. Betts, and A. C. Yee, "Incrementally lossless, broad-bandwidth analog fiber-optic link," in *Proc. IEEE/LEOS Summer Topical Meeting Broad-Band Analog Optoelectron. Devices and Syst.*, Monterey, CA, July 1990, pp. 15–16.

[2] E. Ackerman, D. Kasemset, and S. Wanuga, "An external modulation L-band link with 117-dB spurious-free dynamic range," *Microwave J.*, vol. 34, pp. 158–164, Sept. 1991.

[3] E. Ackerman, S. Wanuga, J. MacDonald, and J. Prince, "Balanced receiver external modulation fiber-optic link architecture with reduced noise figure," in *IEEE MTT-S Symp. Dig.*, Atlanta, GA, June 1993, pp. 615–618.

[4] C. Cox, E. Ackerman, and G. Betts, "Relationship between gain and noise figure of an optical analog link," in *IEEE MTT-S Symp. Dig.*, San Francisco, CA, June 1996, pp. 1551–1554.

[5] C. Cox, G. Betts, and L. Johnson, "An analytic and experimental comparison of direct and external modulation in analog fiber-optic links," *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 501–509, May 1990.

[6] D. Kasemset, E. Ackerman, S. Wanuga, P. Herczfeld, and A. Daryoush, "A comparison of alternative modulation techniques for microwave optical links," *Proc. SPIE*, vol. 1371, pp. 104–114, 1990.

[7] A. Daryoush, E. Ackerman, N. Samant, S. Wanuga, and D. Kasemset, "Interfaces for high-speed fiberoptic links," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 2031–2044, Dec. 1991.

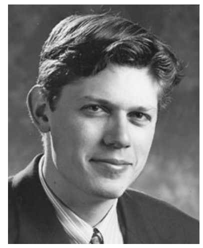
[8] H. Haus *et al.*, "IRE standards on methods of measuring noise in linear two ports, 1959," *Proc. IRE*, vol. 48, pp. 60–68, Jan. 1959.

[9] H. Nyquist, "Thermal agitation of electric charge in conductors," *Phys. Rev.*, vol. 32, pp. 110–113, July 1928.

[10] C. D. Motchenbacher and F. C. Fitchen, *Low-Noise Electronic Design.* New York: Wiley, 1973.

[11] W. Anderson, private communication.

[12] H. Fukui, "Available power gain, noise figure, and noise measure of two-ports and their graphical representation," *IEEE Trans. Circuit Theory*, vol. CT-13, pp. 137–142, June 1966.



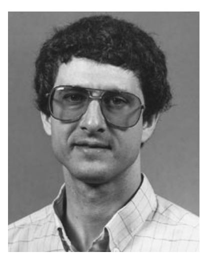
Edward I. Ackerman (M'86) received the B.S. degree in electrical engineering from Lafayette College, Easton, PA, in 1987, and the M.S. and Ph.D. degrees in electrical engineering from Drexel University, Philadelphia, PA, in 1989 and 1994, respectively.

From 1989 to 1994, he was a Microwave Photonics Engineer at Martin Marietta's Electronics Laboratory, Syracuse, NY. He is currently a Technical Staff Member at the MIT Lincoln Laboratory, Lexington, MA, where he develops high-performance analog photonic links for microwave communications and antenna remoting applications.

Charles Cox III (S'70–M'78–SM'95) received the B.S.E.E. and M.S.E.E. degrees from the University of Pennsylvania, Philadelphia, in 1970 and 1972, respectively, and the Sc.D. degree from the Massachusetts Institute of Technology (MIT), Lexington, MA, in 1979.

Since 1979, he has been with the MIT Lincoln Laboratory, where he has been a Technical Staff Member in the Photovoltaic Systems and Applied Photonics Groups. He is currently a Senior Staff Member in the Applied Photonics Section of the Radar Systems Group. For the past 12 years, he has both theoretically and experimentally pursued the development of electro-optic devices and analog fiber-optic links, primarily for antenna remoting.

Dr. Cox is a member of the Optical Society of America and Sigma Xi.



Gary Betts (S'84–M'84) was born in Seattle, WA, on May 14, 1954. He received the B.S. degree in physics from Haverford College, Haverford, PA, in 1976, and the M.S. degree in physics and Ph.D. degree in applied physics from the University of California at San Diego, in 1980 and 1985, respectively.

From 1976 to 1978, he worked at the Westinghouse Electric Corporation, Baltimore, MD, in integrated optics, primarily on the design of geodesic lenses. Since 1985, he has been a Staff Member at the MIT Lincoln Laboratory, Lexington, MA, working primarily on lithium–niobate integrated-optical modulators and analog fiber-optic links. His work at MIT has included microwave-frequency modulators, low-frequency modulators with very high sensitivity, linearized modulators, investigation of optical damage and other drift phenomena in lithium–niobate, and most recently, work on high-power semiconductor optical amplifiers.

Dr. Betts is a member of the Optical Society of America and the IEEE Lasers and Electro-Optics Society (LEOS).

Harold Roussell, photograph and biography not available at the time of publication.

Kevin Ray, photograph and biography not available at the time of publication.

Fred O'Donnell, photograph and biography not available at the time of publication.